It is almost impossible to describe the transmission behavior of a guitar amplifier in its entirety by formulae and diagrams. This is not because the relations and connections would be unknown, but rather because too many dependencies would have to be defined. While the small-signal behavior can easily by specified via the frequency-response, there is – strictly speaking – not even a transfer-function for the large-signal operation because this function is only defined via the LTI-(linear time-invariant)-model. Mixtures of small-signal frequency-response and harmonic-distortion characteristic are either incomplete or too extensive. Non-linear distortion is dependent on frequency and on level, and thus is a bi-variant quantity. There are, in fact, *many* bi-variant quantities: 2nd-order (k_2) and 3rd-order (k_3) harmonic distortion, as well as 2nd-order and 3rd-order difference-tone-distortion, just to name the most important ones. For the – frequently occurring – strong-overdrive condition it is not adequate in the least to assess distortion up to merely the 3rd-order; rather, it would be necessary to determine a multitude of individual harmonic-distortion- and difference-tone-factors, and represent this as a function of two variables. And even if we would make such an effort: the result would be all but impossible to interpret. For example, how would we evaluate a circuit change that results in a reduction of the 3rd-order harmonic distortion at 0,5 V and 1 kHz, while the 2nd-order harmonic distortion increases at 0,8 V and 2 kHz? While at the same time the 4th-order harmonic distortion at 0,8 V (2 kHz) drops strongly but the 2nd-order difference-tone-distortion generally grows stronger? Is this desirable or counter-productive? General judgments such as: *for tubes, 3rd-order distortion (= good) dominates, for transistors 2nd-order (=bad) does* are far too unsophisticated, but unfortunately they keep getting copied again and again from textbook to textbook. Listening experiments remain indispensable. Still, a few fundamental relations can be taken from the theoretical models, after all – even if it the result is not much more than the insight that the circuit layout (than cannot be derived from the schematic) can be highly important, or that tube data have a considerable scatter range. In the following analyses, we will give some data on harmonic distortion for a tube driven via an ohmic source impedance, all the while remaining fully aware that only part of the topic can be covered this way, and that additional research would be highly desirable.

10.1.4 Non-linearity, harmonic-distortion factor

Here is a simple **example** regarding the topic of non-linearity: an amplifier generates – at an input voltage of 1 V – pure 2nd-order harmonic distortion with k_2 = 5%. Let's set its gain factor to v = 1. Now, a second amplifier (also with v = 1) also generating k_2 = 5% at 1 V is connected in series with the first one in a non-reactive fashion. How big would the harmonic distortion of the overall system be?

Would that be: k_2 = 10%, or 7%, or an unchanged 5%?

It is not even possible to answer this question without supplementary data: we do not know the phase of the distortion. In case the 2nd-order distortion is generated in-phase in both amplifiers, k_2 is doubled, but if it happens to be in the opposite phase, the 2nd-order distortions all but cancel themselves out. In both cases an additional 3rd-order distortion appears at k_3 = 0,5%. If there is a random phase-shift between the two amps, k_2 can assume any value between 0 and 10%. Already this simple example shows that it is very difficult to derive any statements about the distortion of the overall system from the non-linear behavior of the single amplifier stages.

So, are you having fun yet, dear audio-engineers? O.K. then – let's go for a **second example**: now both amplifier stages feature pure 3rd-order distortion at k_3 = 5%. Right … use the above: the series connection results in k_3 = 10% for the in-phase condition, and for the out-of-phase condition in k_3 = 0%; plus additionally k_4. Hm … are you sure? Then do turn the page!

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For the pure 3rd-order distortion, the overall system does not distort with k_3 = 10%, but with k_3 = 12,3%, and k_5 = 1% is generated in addition, rather than k_4. Given the anti-phase-condition, we do not see a cancellation but k_3 = 7,5%! Even examples as simple as these show that the results of connections of non-linear systems are rarely understood based on intuition. Moreover, tubes do not exhibit *pure* 2nd-order distortion or *pure* 3rd-order distortion; there will also be distortion of higher order, and in addition the signal will be subject to filter stages – with the result being a highly complex signal processing despite the relatively simple circuitry.

Often, modeling a non-linear circuit starts with the simplification that the system is memory-free. With the investigated system not including any signal memory, the output signal exclusively depends on the input signal at the same instant – with the dependency between both signals described by the **transmission characteristic** (not the transmission function!). This transmission characteristic $y(x)$ is curved (compare to Fig. 10.1.8), but it is time-invariant and excludes any hysteresis. The characteristic may be expanded into a power-series (Taylor/MacLaurin) around the operating point – the smaller the drive levels, the more precisely this works. Put in another way: the more the amp is driven, the less the power series is appropriate. This is easily understood: a limiting characteristic has two horizontal asymptotes, which is incompatible with a power-series converging towards infinity. In this situation, wouldn't the arctan-function seem to be a much better starting point? Yes, indeed – but it would be one that is far from intuitively accessible: how is e.g. $x = \hat{x} \cdot \sin\omega t$ mapped again onto y? With $y = \arctan(\hat{x} \cdot \sin\omega t)$. O.k., I see. So how does the harmonic distortion depend on the drive levels? Well, we would have to develop a series-expansion of the arctan … Phewww – that means we might as well expand the transmission characteristic into a series:

$$y = a_0 + a_1 \cdot x + a_2 \cdot x^2 + a_3 \cdot x^3 + ...$$ Series expansion of the transmission characteristic

In this expansion, a_0 is the DC-offset that is separated in most circuits by the coupling capacitors; we ignore this offset. a_1 is the gain – for an input tube this might be e.g. -54. Now we get to the non-linearity: using, for example, the pure 2nd-order distortion (i.e. $a_i = 0$ for $i > 2$), we obtain

$$y = a_0 + a_1 \cdot \sin\omega t + a_2 \cdot (\sin\omega t)^2 = a_0 + a_1 \cdot \sin\omega t + a_2 \cdot (1 - \cos 2\omega t)/2$$

Due to the non-linearity, the DC-component has changed but we can again ignore it. There is now a new spectral line at twice the frequency. The ratio of the RMS-values $k_2 = \tilde{y}_2 / \tilde{y}$ is designated the **2nd-order harmonic distortion** k_2. $\tilde{y}_2$ is the RMS-value of the 2nd-order harmonic (at 2ω) and $\tilde{y}$ is the RMS-value of y. Let us set, as an example, a_1 = 1 and a_2 = 0.1 – this yields $k_2 \approx a_2 / 2a_1 = 5\%$. Connecting two such systems in series, a series-transformation $z(y(x))$ is the result:

$$z = a_0 + a_1 \cdot y + a_2 \cdot y^2 = a_0 + a_1\left(a_0 + a_1 \cdot x + a_2 \cdot x^2\right) + a_2\left(a_0 + a_1 \cdot x + a_2 \cdot x^2\right)^2$$

Assuming again $x = \sin(\omega t)$, the amplitudes (or rather the RMS-values) of the individual harmonics can be calculated. What is striking in view of the second bracket of the equation is that the offset (a_0) now not only influences the DC-component but the 1st and 2nd order harmonic, as well! Moreover, we notice the generation of a 4th-order harmonic due to x^4 – although its amplitude is so small that it may be disregarded. From x^2, a DC-component and the 2nd-order harmonic result, and from x^3 we derive the 1st-order and the 3rd-order harmonics. x^4 generates a DC-component plus the 2nd-order and 4th-order harmonics. So, everything depends on everything else, more or less.

In summary: for $a_0 = 0$, the levels of the 1st, 2nd, and 3rd harmonic depend on a_1 and a_2, and only the 4th harmonic depends solely on a_2. A simplification yields:

$$z = x + 2a_2 \cdot x^2 + 2a_2^2 \cdot x^3 + a_2^3 \cdot x^4; \qquad \text{for } a_0 = 0 \text{ and } a_1 = 1.$$

Neglecting higher-order effects, we can state that the series connection indeed leads to double the 2nd-order harmonic distortion. Also, if one system is set to be $y = a_1 \cdot x + a_2 \cdot x^2$ and the other is set to be $z = a_1 \cdot y - a_2 \cdot y^2$, k_2 can be approximately fully compensated. The simplifications we introduce here are purposeful – they do not generate any large errors.

For the purely 3rd-order distortion, the mapping is: $y = x + a_3 \cdot x^3$. The offset a_0 is set to zero in this case, and the linear term is set to 1 ($a_1 = 1$) as a simplification. The series connection yields:

$$z = y + a_3 \cdot y^3 = x + a_3 \cdot x^3 + a_3 \cdot \left(x + a_3 \cdot x^3\right)^3 = x + 2a_3 \cdot x^3 + 3a_3^2 \cdot x^5 + 3a_3^3 \cdot x^7 + a_3^4 \cdot x^9$$

Again, we could disregard all higher-order terms and assume that the 3rd-order harmonic distortion will be doubled. However, the 1st and the 2nd harmonic are dependent on all summands, and the resulting effect is not at all that minor:

$$\sin^3(\varphi) = \frac{1}{4}\big(3\sin(\varphi) - \sin(3\varphi)\big), \qquad \sin^5(\varphi) = \frac{1}{16}\big(10\sin(\varphi) - 5\sin(3\varphi) + \sin(5\varphi)\big)$$

The summation of all terms of the expansion has the effect that the RMS-value of the 3rd harmonic is not only doubled but rises by a factor of 3,7. At the same time, the RMS-value of the overall signal increases by half, yielding k_3 = 12,3%. If the sign of a^3 in one of the two systems is inverted, x^3 can be reduced to zero but the remaining members of the series deliver a significant contribution to the 3rd harmonic. The amplitude of the latter therefore does not go down to zero but decreases merely to 7,5%.

If the **offset** (a_0) is not set to zero, the situation becomes even more complicated. The same happens if we do not keep the limitation on purely 2nd- and 3rd-order distortion, respectively. So: with two nonlinear systems connected in series, *both* generate 2nd- as well as 3rd-order distortion.

$$z = b_0 + b_1 \cdot y + b_2 \cdot y^2 + b_3 \cdot y^3; \qquad y = a_0 + a_1 \cdot x + a_2 \cdot x^2 + a_3 \cdot x^3$$

O.k. – computing this is not impossible; we could multiply that out. The added value would not be that big, however. It is already clear now that the RMS-value of each harmonic will be dependent on many coefficients. Also, there will be cancellations of components if there are opposing algebraic signs. These cancellations will be drive-level-dependent, though – or at least there will be drive-level-dependent maxima and minima. The individual harmonic distortion components will not simply experience a monotonous increase with rising drive-levels but can pass through complicated curves. The individual system may generate exclusively even-order distortion (k_2, k_4, ...), but the series connection of two such systems may still show a predominant odd-order distortion. If we now consider that even simple guitar amplifiers do not contain one but four tube-stages, and that each tube introduces distortion both at its input *and* at its output, and that moreover tone-filters will change amplitudes and phases … this is where we start to catch a glimpse of how complex a guitar amp in fact is.

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The following figures present the dependency of individual harmonic distortion components on the input signal level. N.B.: Another way of quantifying distortion is expressing how much lower the level of the distortion is compared to the original (undistorted) signal. This approach yields the so-call harmonic **distortion attenuation** a_{ki} calculated from the harmonic distortion factor k_i (as we have been using it so far) as:

$$a_{ki} = 20 \cdot \lg(1/k_i)\ \mathrm{dB} \qquad \text{for e.g. } k_2 = 5\% \;\rightarrow\; a_{k2} = 26\ \mathrm{dB}.$$

All measurements were done with a regularly heated ECC83 with the intrinsic distortion factor of the analyzer being negligible (k < 0,001%, CORTEX CF-100). The cathode of the tube was connected to ground via 1,5 kΩ // 25 µF (Fig. 10.1.1), and the plate to U_B via 100 kΩ. To model the load, the plate was additionally connected to ground via a 0,33-µF-100-kΩ-series-circuit. The signal was fed to the grid from a low-impedance generator (CORTEX CF-90) via the grid-resistor R_g. For one row of measurements, R_g was 33 kΩ (corresponding to a classic tube amp scenario that is fed from a low impedance source), and for the other row it was R_g = 133 kΩ (corresponding the additional source impedance of 100 kΩ as it can be present if a guitar with a passive pickup is operated around its resonance frequency, compare to Fig. 10.1.10). The supply-voltage U_B amounted to 200 V and 250 V, respectively, i.e. typical settings for input stages (**Fig. 10.1.12**).

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Abb. 10.1.12: Distortion attenuation as function of the generator level, R_g and U_B vary. 0 dBV ≙ 1V_{eff}.
These graphs are reserved for the printed version of this book.

While the a_{k3}–curve maintains its shape and predominantly experiences "merely" a shift, the minima and maxima of the 2nd-order distortion change rather drastically. *So will you tell me how that sounds, already?* would be an obvious question … however: nobody actually listens to the plate-voltage of the preamp-tube, and therefore the *sound* of that signal is irrelevant. Highly relevant would be how the differences mentioned above affect the loudspeaker voltage, but this would require the consideration of a myriad of additional parameters and go beyond the constraints give here. Unfortunately.

Another question relates to the **tube**: RCA, Tungsram, Telefunken, Chinese, Russian, NOS, little/much used, and whatever other difference there might be? Simple answer: the tube came out of the box that served here as container for tubes since 1965, and was re-stocked many times since. An ECC83 cost DM 7,50 (about € 3,25) in Germany in 1965; today is offered for € 6.-. It could also set you back € 25, or even more, though. Without a doubt, tubes of the same type can differ a lot – the label "ECC83" does not indicate any special sound. Selection processes performed by the supplier *may* be helpful but do not *have* to be. Pricy tubes are not necessarily in principle better than cheap ones; in particular, "NOS" (i.e. the tube that has spent 50 years on the shelf without being touched) does not guarantee a "super-sound".

In **Fig. 10.1.13** we see differences that can occur when we change tubes (all measurements taken with ECC83s). A tube was simply unplugged and another was plugged in, instead. It is intentional that manufacturers are not identified here, since we do not have a representative sample. We did not investigate whether an old 80-$-NOS-tube delivers similar or entirely different curves – confronted with its measurements, it might have experiences a kind of final deadly shock. Plus, strictly off the record: for the analyst, this is somewhat like the situation experienced by Galileo's colleagues who did not even want to look through the telescope to see Jupiter's moons – some of us in fact don't really want to know.

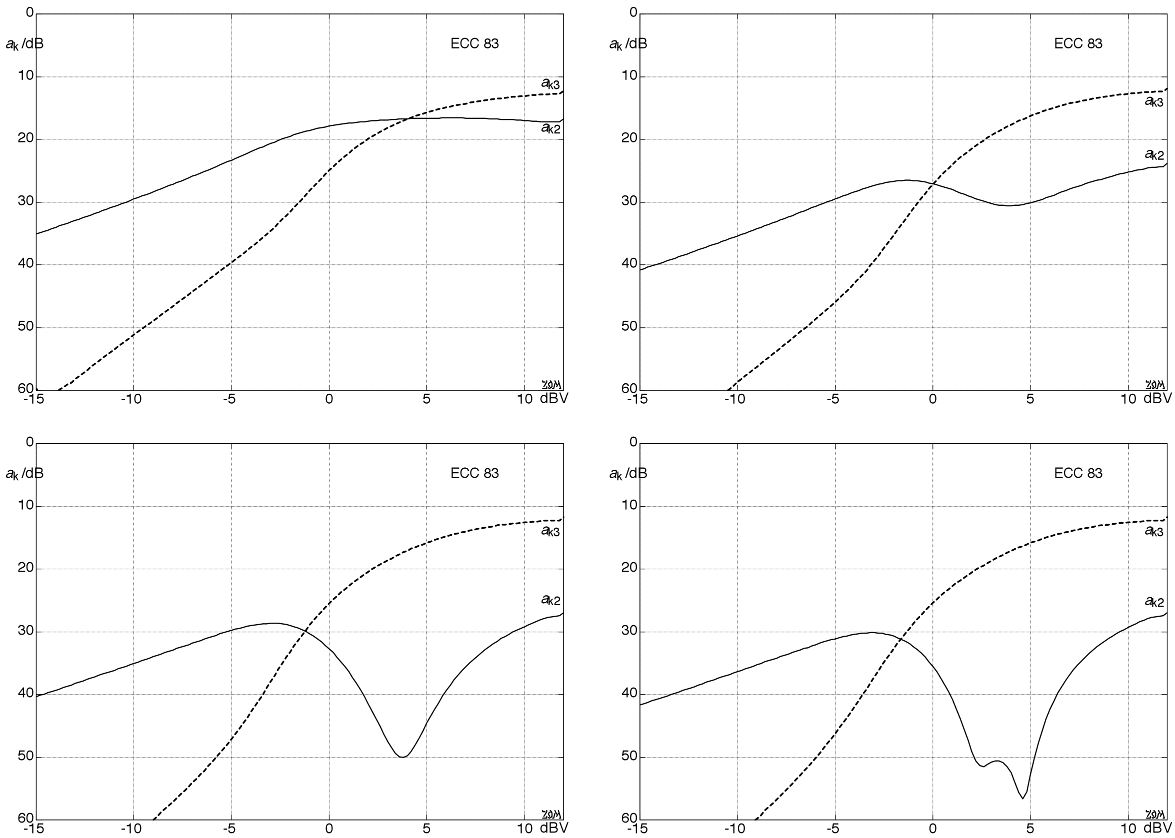


Fig. 10.1.13: Differences in harmonic distortion attenuation caused by swapping tubes. U_B = 250V, R_g = 33kΩ.

Now back *for* the record: already at an input voltage of 300 mV$_{eff}$, the harmonic distortion in the input stage of a guitar amp can reach 3%. For small input voltages, 2nd-order distortion is predominant while from 0,25...1 V, 3rd-order distortion dominates. The location of the border between the two distortion types depends on the grid-resistor, on the supply-voltage, and on the ECC83-specimen. The distortion is not inherently unwelcome but rather typical for a guitar amp of this construction.

So far, we have varied, as parameters, the grid-resistor, the supply-voltage and the tube itself. As the **supply-voltage** changes, the plate-current, the plate-voltage, and the grid/cathode-voltage change, as well. Of course, more parameters vary – but right now we look only at these three. For example: increasing the supply-voltage from 200 V to 250 V increases the plate-voltage from 131 V to 165 V, and U_{gk} decreases from -0,97 V to -1,23 V. Another method to vary the anode current is the so-called “cathode clamp”: here, the cathode-voltage is imprinted (i.e. kept constant) using a separate power supply. One could think that the cathode-voltage could not change anyway due to the capacitor connected in parallel – but in fact, it can: a 2^{nd}-order-distorted sine tone will generate a DC-component ($f = 0$) that shifts the operating point. The following figures show the effects of a relatively small change in the grid/cathode-voltage on the distortion /(**Fig. 10.1.14**).

Abb. 10.1.14: Harmonic distortion-attenuation dependent on the input level with varying grid/cathode-voltage. These figures are reserved for the printed version of this book.

It is clearly visible that even apparently minor changes in the operating point have considerable effects on the non-linear distortion. For reasons of clarity, no higher-order distortion products are included in the figures; it can be stated, however, that they are highly similar. The operating point of the tube is far from fixed but drifts while the amp is being played. One cause for this is found in the non-linearities already mentioned, and another lies with the time-variant supply-voltage. The latter depends on the plate-currents of the power stage and the internal impedance of the power supply and will change depending on the output power of the amp at the given moment (see Chapter 10.1.6). For the Fender Deluxe we investigated, this variation was as much as between 210 and 247 V, after all . . .

Now, what is so special in a tube amplifier compared to other amps? Looking at the preamplifier, there are differences in particular in the non-linear behavior. There are, in addition, compressor effects and linear filtering – this will be elaborated upon a bit later. The **operational amplifier** (OP) appears to be a modern alternative to the tube. It has an operational range to above 1 MHz and its harmonic distortion may be reduced to 0,001%. These are, however, all properties that a guitar amplifier should not actually have! An OP may only be considered as an alternative if additional circuitry simulates the non-linear behavior of the tube. That this is not entirely trivial was shown in the preceding paragraphs. **Fig. 10.1.15** depicts the drive-level-dependent increase of the distortion for the ideal OP in comparison with to the tube. The hard amplitude limiting ("clipping") leads to a steep distortion increase that is atypical for a tube.

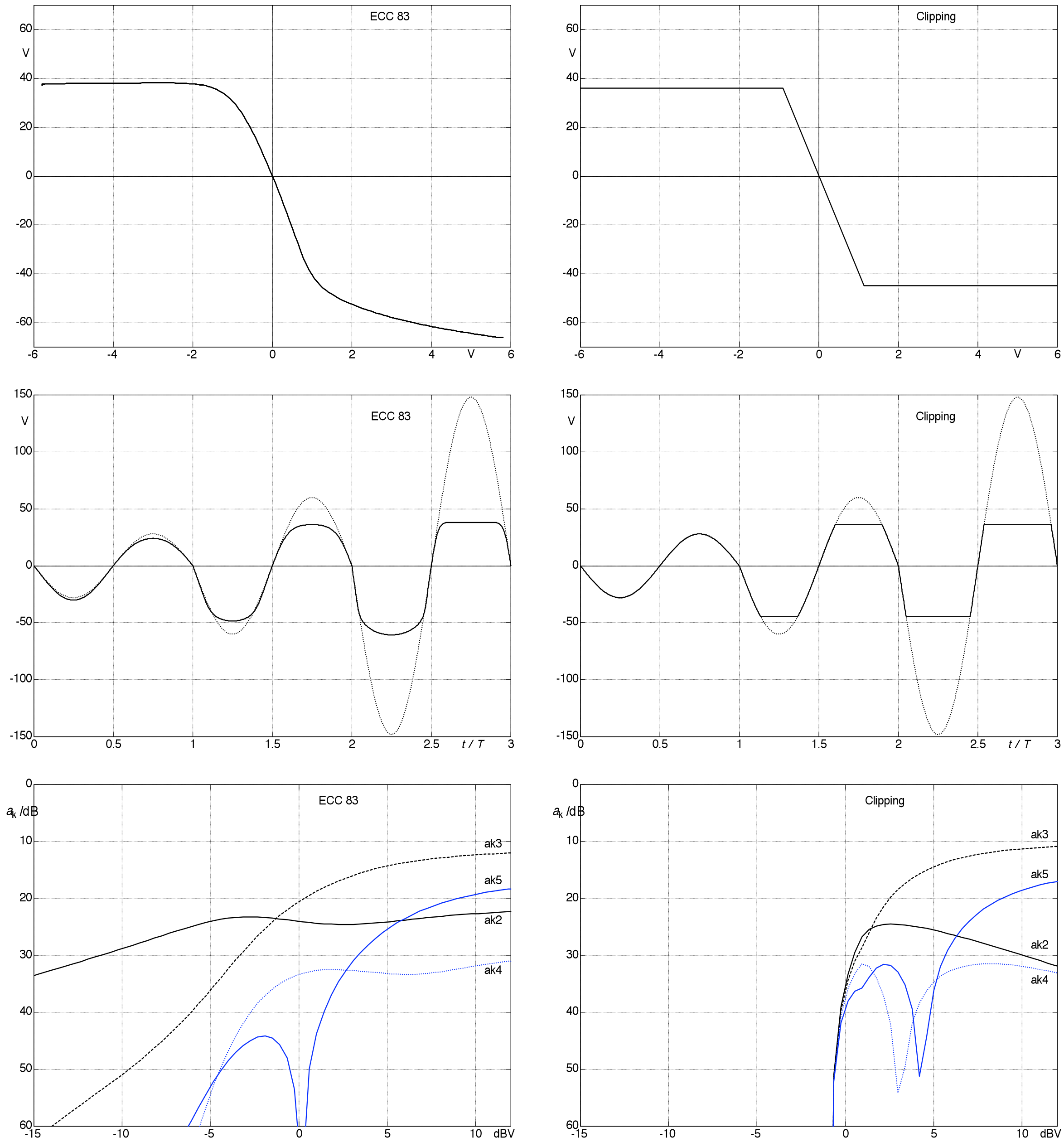


Fig. 10.1.15: Distortion for tube-typical limiting (left) and hard OP-clipping (right). By the way: the designation "ideal OP" does not imply that the OP would be ideal for playing guitar through it. NB: The OP-offset was adjusted for an asymmetry similar to a tube.

The rise of the distortion will be considerably flatter if the signal limiting is realized not by the OP itself but by two silicon diodes (1N4148) in an anti-parallel connection (**Fig. 10.1.16**). If the two measured diodes were perfectly identical, only odd-numbered distortion products would appear; due to small production-related differences, we also obtain even-numbered distortion products in this example. The 3rd-order distortion of this diode circuit already shows strong similarities to the triode circuits measured in Fig. 10.1.15 but the 2nd-order distortion is not reproduced yet. It is not very demanding to design – using a combination of germanium and silicon diodes – a non-linear two-pole the distortion behavior of which sounds similar to that of a tube. The exact reproduction of tube-distortion is not even required for this; an approximate modeling suffices.

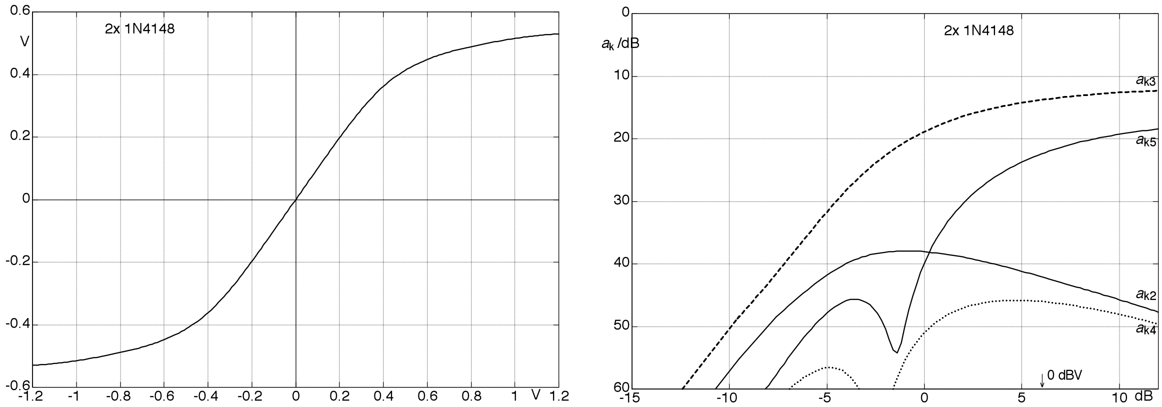


Fig. 10.1.16: Signal limiting using two anti-parallel silicon diodes (1N4148) fed from a 20-kΩ-resistor. The level reference on the abscissa of the right-hand picture is chosen to match the representation in Fig. 10.1.15.

It is not only the harmonic distortion that is different in tube and OP, but the **compression** is, as well (**Fig. 10.1.17**). This difference is not big, but may be compared to the so-called "sagging" – a modulation caused by the power supply (Chapter 10.1.6). In the attack phase of a tone, a tube amp may lend that extra little bit of power that can be decisive when competing with other instruments. That tube amplifiers can be louder than transistor amps rated at the same power is due in particular to the higher output impedance, but may also have to do with the weaker compression (= increased dynamics). Of course, this is not generated by the preamp-tube alone but by the overall circuit.

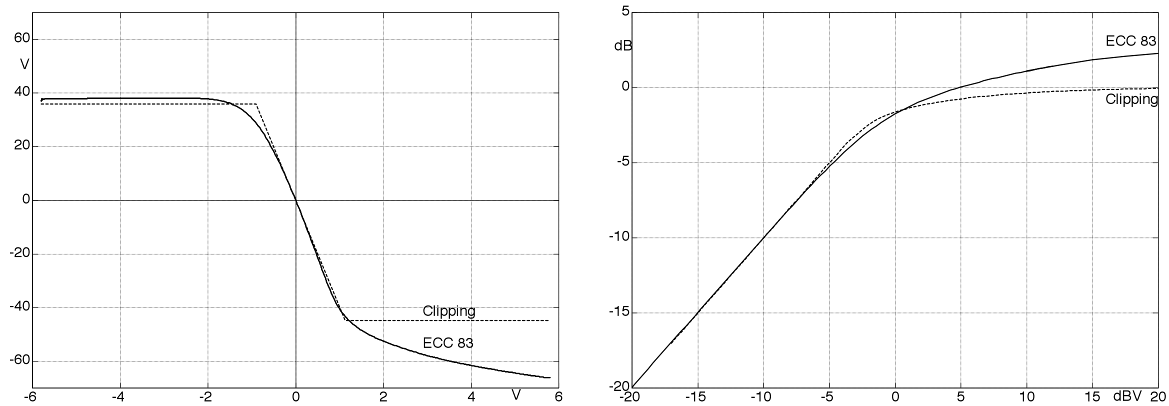


Fig. 10.1.17: Signal-limiting in a tube and an ideal OP. Equal small-signal gain.

We will dedicate a later chapter to the power-delivery, but at this point a fundamental aspect of the connection of non-linear systems may already be briefly introduced: the transformation caused by the individual systems are generally not **commutative**, i.e. the individual systems cannot be simply interchanged in their sequence. For this reason, it is not possible to replace an amplifier consisting of a plurality of non-linear and linear systems by a single non-linear stage and a single filter-stage. Special consideration needs to be given to the fact that already the coupling capacitor that taps the signal from the plate is such a filter-system, even if the associated cutoff frequency of this high-pass is very low.

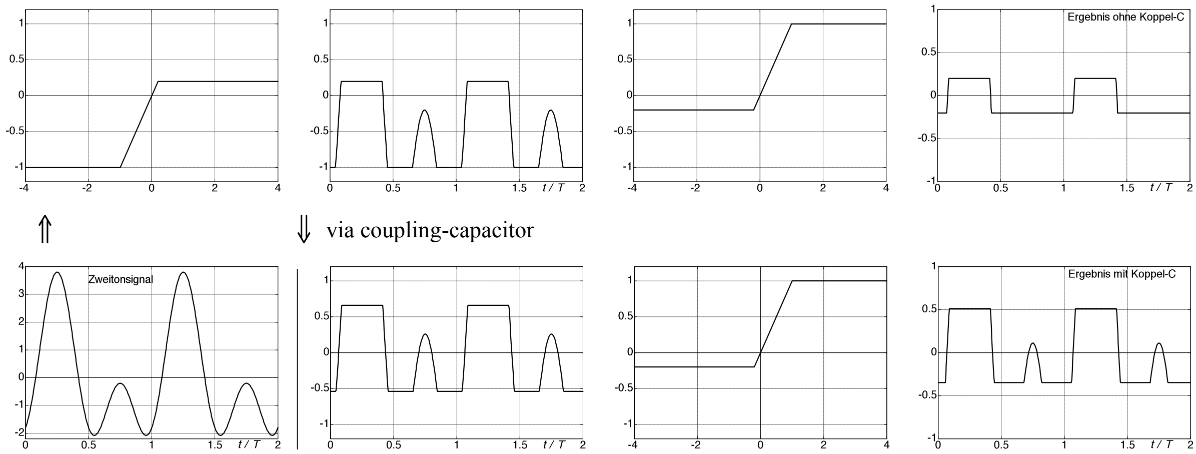
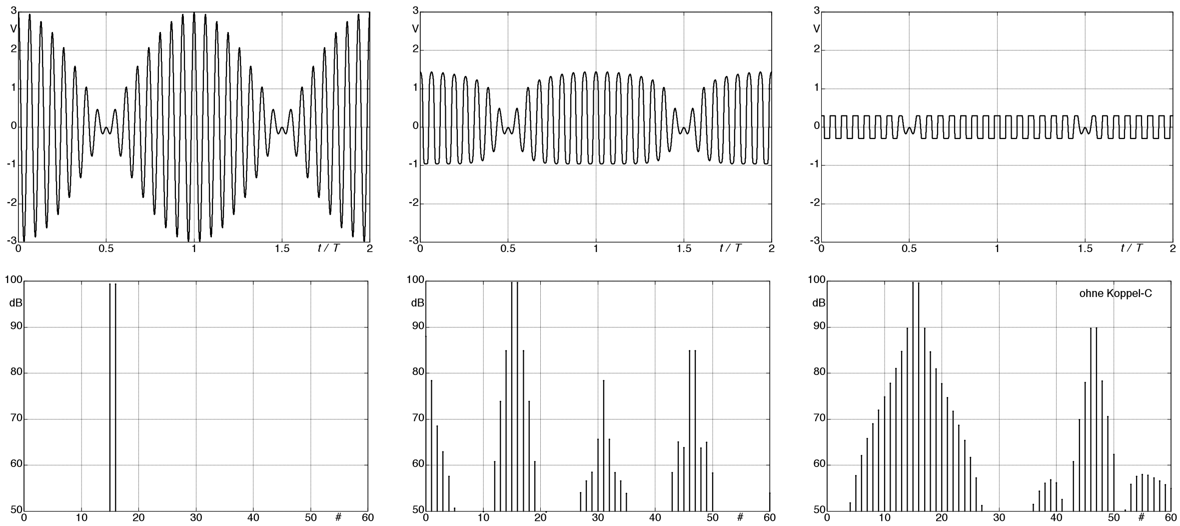


Fig. 10.1.18: Half-wave limiting in a sequence of systems; top: w/out coupling cap, bottom: with coupling cap.

Fig. 10.1.18 shows an example: a two-tone signal first passes a stage limiting the positive half-wave and then a second stage limiting the negative half wave. If these two stages directly follow each other, the result is a signal limiting on both sides as depicted in the upper row of pictures. However, if a **coupling capacitor** is connected in between the two limiting stages, we obtain an entirely different output signal (lower right picture). With the coupling capacitor connected *ahead of* the first limiter stage, it would have no effect because the two-tone signal is already without any DC-component. The same result would be obtained with the capacitor positioned *after* the two limiting stages. However, connected *between* the two stages, the capacitor will change the signal even if the cutoff frequency is far below the two frequencies contained in the two-tone signal. Now, let's put this in the context of a guitar amplifier: since the plate-voltage is 150 – 200 V even without any drive signal, a coupling capacitor is required to split off the AC signal. Together with the input impedance of the subsequent stage, this capacitor forms a high-pass. In many circuits, its cutoff frequency is so low that it does not seem to have an effect. For example, in the Fender Bassman (held in highest regard also by guitarists), we find f_g = 3Hz (50nF/1MΩ) which is way below any normal frequency found in the guitar. However, Fig. 10.1.18 shows that this coupling cap has an effect despite its low cutoff frequency: the non-linearity will generate extremely low frequencies (0 Hz if you wait long enough …) that are split off by the high-pass. Taken by themselves, these low frequencies would be inaudible. However, they do determine the position of the operating point and therefore influence the distortion of the subsequent stage. The specific value of the cutoff frequency also has a significance because it determines how fast the transient processes run (Chapter 10.1.6). This example very clearly shows that design rules valid for linear operation can lose their relevance in an overdrive scenario.

In our second example (**Fig. 10.1.19**), a signal of two sine signals close in frequency and beating against each other is first distorted with a tube characteristic, and then subject to hard clipping. The scaling of the ordinate is chosen for all pictures such that equal amplitudes result for small-signal operation. Given such an extreme clipping one could surmise that the “soft” tube distortion occurring first would not have any actual effect since subsequently we have clipping, anyway. As long as there is no coupling capacitor between the two distortion stages, this assumption is indeed correct. However, as a coupling capacitor is introduced, the signal changes – in particular in the low-frequency region and in the area of the summation frequency of the two sinuses (in this example around the 31st harmonic).



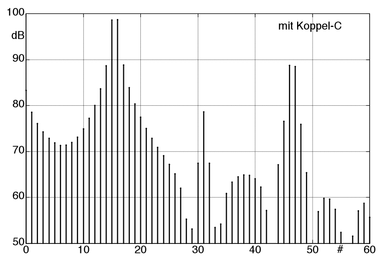


Fig. 10.1.19: Beat-signal (1st column).
Distorted with tube characteristic (2nd column),
then subject to hard clipping (3rd column).
The spectrum given in the 3rd line shows
the effect of the coupling capacitor.

To *round* off this section, let us bring the “round” tube distortions face to face with the typical OP-clipping. If we trust literature, then the latter is the reason for the “harsh” transistor-sound – as opposed to the soft tube sound. Sure, there are differences in the spectrum (**Fig. 10.1.20**), but in fact we also find similarities. In any case, the visual impression (“a round signal shape will sound more round, as well”) should not be overrated; tube- and transistor-amps differ in much more than just the rounding of the signal-shapes. Only the connection of several systems makes for the amp. Or, rather, for the sound …

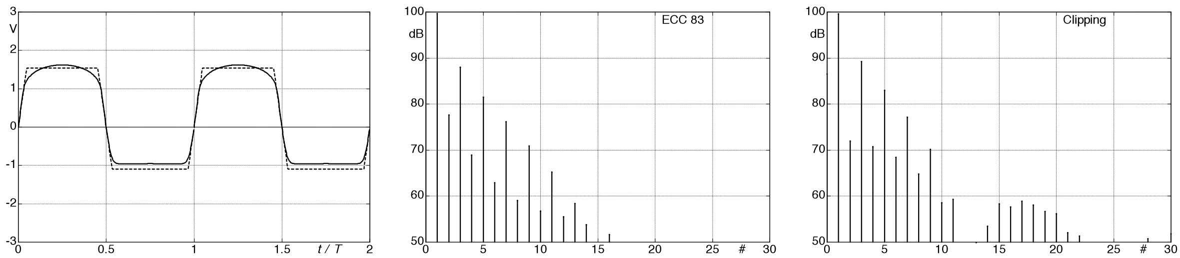


Fig. 10.1.20: Tube distortion (ECC83) compared to hard OP-clipping, driven by a sinusoidal signal; the distortion levels below 60 dB correspond to a harmonic distortion of < 1% in this example.

Special consideration needs to be given to slew-rate-limiting since such non-linearity does not occur in tube amplifiers. The **slew-rate** *SR* is the speed of change in the signal i.e. the derivative *dU/dt*, usually given in V/µs. For a sinusoidal signal, the maximum slew-rate is present at the zero-crossing: $SR = 2\pi f \cdot \hat{u}$. A voltage amplitude of 13V (a typical OP amplitude) results in a slew-rate of just about 1 V/µs at f = 12 kHz. If the maximum slew-rate that the amplifier can provide is smaller than the signal slew-rate, then non-linear distortion results. In contrast to a low-pass (the *linear* transformation of which can alternatively be specified by a cutoff-frequency and a time-constant), the slew-rate-limiting is a *non-linear* transformation that changes the signal shape in particular close to the zero-crossing (**Fig. 10.1.21**).

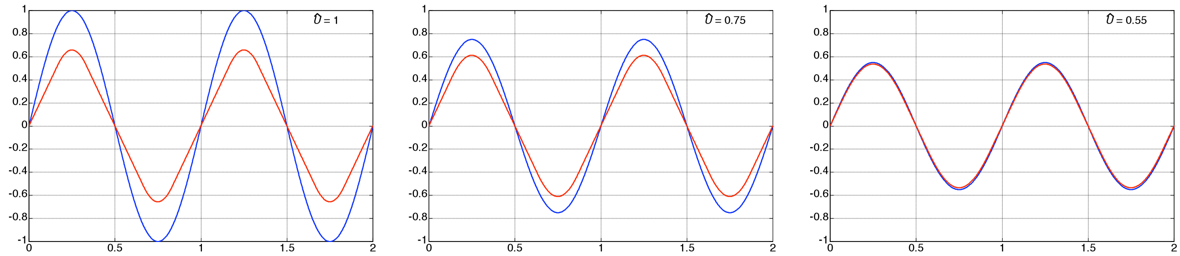


Fig. 10.1.21: Sine-functions of different amplitude (——), non-linear transformation w/slew-rate-limiting (——).

Although in principle the slew-rate may be limited for rising signals to a different value compared to falling signals, both values are almost equal for most operational amplifiers: for example for the (outdated) LM-741: SR_{max} = 0,5 V/µs, or for the TL-071: SR_{max} = 13 V/µs. With a SR_{max} = 0,5 V/µs, the maximum frequency for distortion-free, full-drive-level operation is only 6 kHz. One could assume that this would suffice for a guitar amp since most magnetic pickups limit their spectrum at the most at this frequency. However, this assumption overlooks the possibility of overdrive: if this 6-kHz-tone overdrives the OP by a factor of 10, then the signal-slew-rate is also 10 times as quick at 5 V/µs. **Fig. 10.1.22** shows that slew-rate limiting and clipping are two different kinds of non-linearity: clipping limits the too-large values of the signal while slew-rate limiting confines the value of the slope of the signal. If both types of distortion happen in one and the same stage, the sequence needs to be considered: the two transformations are not commutative!

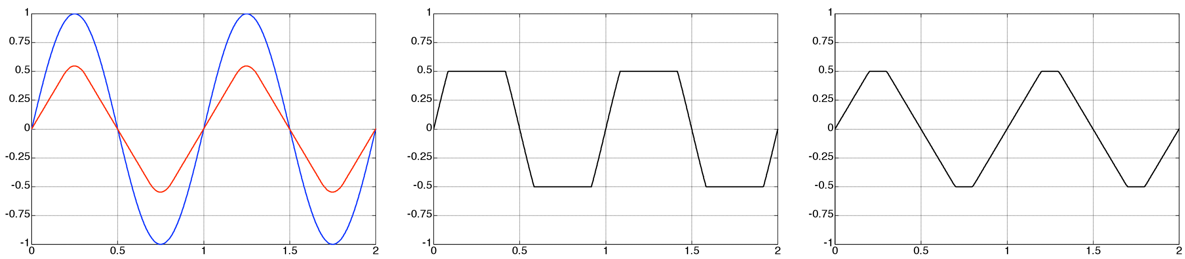


Fig. 10.1.22: Sinusoidal signal (——), slew-rate limiting (—— left). Sinusoidal signal with clipping (middle). Sinusoidal signal with slew-rate limiting and subsequent clipping (right).

The principles of circuit design are the reason that we get slew-rate limiting with an OP but not with a tube (in a comparable manner, anyway). In the OP a number of subsequent stages generate a very high amplification (e.g. 100.000) that is then reigned in by negative feedback to e.g. 50. The same gain is accomplished in tube amps in a single stage without or with very little negative feedback. The high gain of the OP forces another difference: in order for the overall feedback-loop to remain stable, a low-pass characteristic (e.g. with a cutoff at about 100 Hz) is required in the forward branch (i.e. in the pure OP without the feedback network).

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Working in the framework of a model, we may replace the typical OP by the building blocks shown in **Fig. 10.1.23**: a comparator (subtractor) is followed by a first amplifier, a first limiter, a 1st-order low-pass (with e.g. the aforementioned 100 Hz cutoff), a second amplifier, a second limiter. A (negative) feedback branch connects the output to the other input of the subtractor). The DC-gain is e.g. 100.000; the gain drops off with 1/*f* from 100 Hz, reaching the value of 1 at 10 MHz (the so-called transit frequency).

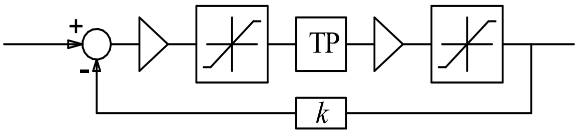

Fig. 10.1.23: Block-diagram of a typical operational amplifier with negative feedback.

Limiting can occur at *two* places in this amplifier with different effects on the external world: limiting in the output stage will introduce clipping, while limiting occurring ahead of the low-pass will introduce slew-rate distortion. We may see the low-pass approximately as an integrator – this will give us an easily understandable model for the limiting of signal rise-times. If the amplifier is driven with a low-frequency signal, the output stage will limit first and we get clipping. With a high-frequency signal, the stage ahead of the low-pass will limit first and slew-rate-type limiting happens.

It really gets interesting for a mixture of tones, e.g. with a two-tone signal consisting of a 1st and a 2nd harmonic (**Fig. 10.1.24**). This signal has the same peak value both on the negative and the positive side and would be symmetrically limited given a point-symmetric limiter-characteristic. However, since the zero-crossings have slopes of different steepness, the slew-rate distortion has a different effect on the two half-waves, resulting in a shifting of the signal: it moves away from the zero-line and becomes asymmetric. In the OP, the negative feedback would immediately become active (the loop gain is indeed very high at low frequencies), and a counteractive offset voltage would result, with the slew-rate limited signal losing its quality of being DC-free, and experiencing a shift towards the negative (middle picture). Now the clipping is added that predominantly limits the negative half-wave – despite the fact that the original two-tone signal is in fact symmetric with regard to the horizontal axis. The processing of the second signal – a superposition of 1st and 3rd harmonic (right-hand picture) – is just as interesting: the slew-rate distortion does not only reduce the signal but distorts it in a non-linear fashion. Still, dents remain at the location of the extreme signal values – in contrast to the effect of pure clipping. These examples show that the slew-rate distortion occurring in OP-circuits has a very different effect compared to pure clipping process that is often seen as the sole reason for distortion. In the typical operational amplifier, slew-rate distortion does, however, not appear by itself but always in combination with clipping.

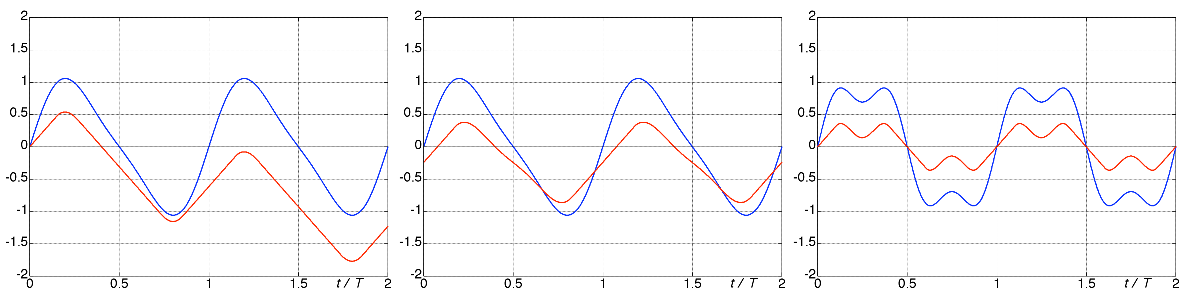

Fig. 10.1.24: Slew-rate limiting: left and middle: 1st and 2nd harmonic. Right: 1st and 3rd harmonic. Two-tone signal (——), slew-rate-limited signal (——).

The maximum slew-rate that an OP can handle may vary dramatically – depending on the OP-type: in this respect the old **µA709** (one of the first universally usable operational amplifiers) was particularly inept. Its maximum slew-rate was (at 0,25 V/µs) so small that full drive was only possible up to 3 kHz at most. Since at the time of the introduction of this OP (1966), harmonic-distortion measurements were usually carried out only at 1 kHz, the slew-rate distortion often remained undiscovered. The **µA741** introduced two years later was able to deal with double the slew-rate but that was still not enough: 10-fold overdrive with a 3-kHz-signal requires 2,5 V/µs. Only later OP-amps – such as the **TL071** at 13 V/µs – reach faster regions. By the way: what is in fact the typical slew-rate of the voltage generated by a magnetic pickup? Of course, this depends on many parameters; in Fig. 10.1.5, for example, 0,06 V/µs is reached. Feeding this signal to a **Music-Man**♣ guitar amplifier, it will be amplified 20-fold in the first OP. To avoid any slew-rate distortion, the OP would have to be able to take on 1,2 V/µs. However, the LM1458 used in some Music-Man amps cannot go beyond 0,5 V/µs without distortion (just like the LM307H used as an alternative). Not all Music-Man amplifiers used these slow LM1458 or LM307H in their input-circuits: some work with the fast TL071 (13 V/µs) … but then feed the signal to a LM1458 in the third amplification stage. Worse: for the input-OP, the gain in the treble range can be increased from 20 to 120 via the “Bright”-switch, increasing the necessary slew-rate value by another factor of six. The distortion generated by this is therefore tube-**un**typical. That the Music-Man amp has a tube power amp ahead of the loudspeaker will therefore not guarantee the same sound compared to an amplifier working exclusively with tubes in its signal path.

Of course, tubes are not infinitely fast, either; however in most cases in tube circuits the rise-time is already limited in the grid-circuit via a low-pass. While this low-pass is non-linear due to its (Miller-) capacitance depending on the voltage-gain of the tube, this non-linearity has an entirely different effect compared to slew-rate limiting.

The following table lists the slew-rate values for some operational amplifiers. Depending on the manufacturer, the numbers differ somewhat: for the LF356, for example, we find both 10 V/µs and 13 V/µs. The first letters in the designation may indicate the manufacturer (e.g. LM 741, or SG 741, or µA 741), while the last letters specify housing types, or temperature ranges, or amplifiers with selected data (e.g. LM 307 and LM 307H). These supplementary letters are, however, not standardized but specific to the respective manufacturer. For some types, the open-loop gain (and thus the slew-rate) can be changed via an externally connected capacitor (so-called compensation, e.g. in the LM 301A).

35 V/µs: HA 5147, OPA 404,
13 V/µs: TL 071, LF 351, LF 353, LF 356,
10 V/µs: LM 302, LM 301A (uncompensated),
6 V/µs: NE 5534, LF 355,
0,5 V/µs: LM 107, LM 207, LM 307, LM 741, µA 748, RC 1458, RM 1558,
0,2 V/µs: OP 07,
0,1 V/µs: LM 108, LM 208, LM 308 (each compensated),

Table: Slew-rates of some selected operational amplifiers (guide values).

♣ Amplifiers and instruments, founded in 1972 by Leo Fender (and Tom Walker), sold to Ernie Ball in 1980.

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